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A TRANSISTORIZED COMMUTATOR

A THESIS

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by

Spurgeon L. Robinette

In Partial Fulfillment

of the Requirements for the Degree

Master of Science in Electrical Engineering

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A TRANSISTORIZED COMMUTATOR

APPROVED

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ACKNOWLEDGMENT

The author wishes to express his appreciation to Dr. B. J. Dasher, director of the School of Electrical Engineering, who suggested the topic for this thesis, and to Mr. H. L. McKinley and Mr. W. B. Warren who gave much helpful advice and encouragement during the work.

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SUMMARY

It often becomes necessary to transmit information derived from several input circuits or channels over a single output channel. The device which effects this is called a commutator. In it, the voltages in the input circuits are sequentially sampled, and the samples are then presented, in time sequence, in the single output circuit.

The Electrical Engineering School of the Georgia Institute of Technology desired an eight-channel transistorized commutator for two reasons. It would aid in deciding the feasibility of a larger transistorized commutator for use as a part of speech analysis equipment. It would also serve as a classroom teaching aid which would permit simultaneous oscilloscopic display of up to eight independent voltages.

The development and construction of the commutator was undertaken as the problem for this thesis. The resulting device is a matrix of thirty-two point-contact diodes, arranged so that each of the eight matrix inputs is connected in turn to the output by means of properly scheduled voltage combinations in three control-circuit pairs. The voltages needed in the control circuits are provided by a chain of three bistable transistor multivibrators.

A square wave of voltage (from an external generator) is required to synchronize the multivibrator chain. A negative going voltage pulse is derived from each cycle of the square wave, and each pulse causes the first multivibrator to change state, so that it runs at half the

synchronizing frequency. In like manner a pulse train from the first multivibrator triggers the second, and pulses from the second trigger the third, with frequency halving in successive stages. The three multivibrators thus run at 1/2, 1/4, and 1/8 the synchronizing signal frequency. This mode of synchronization of the commutator has proved stable in operation.

The multivibrators are designed to operate from cutoff to a point short of saturation. Switching speed is thus kept high.

In addition to the diode matrix and the multivibrators, there is included in the commutator a cascaded, two-transistor emitter follower (Darlington circuit) which provides a low output impedance. A fast pulse stretcher was also developed to improve the oscilloscopic presentation, but it is needed only at the highest sampling rate of 150,000 samples per second.

Satisfactory oscilloscope waveforms are obtained without the pulse stretcher with sampling rates from below 10,000 samples per second to 100,000 samples per second. Visible switching effects (pedestals) are 2.5 to 3 microseconds in duration.

Requirements for input conditions for the commutator are:

1. The signal must be negative with respect to ground.
2. It must not exceed three volts in magnitude.
3. The source impedance of the signal must be small with respect to 10,000 ohms, e.g., 100 ohms.
4. The Nyquist sampling criterion must be met, i.e., the signal spectrum must include only negligible energy at frequencies above half the channel sampling rate.

If the above restrictions are met, the output voltage will vary linearly with input voltage, over a 35-decibel dynamic range.

A positive 9-volt and a negative 6-volt dry cell constitute the power supply. About 24 milliamperes is drawn from the batteries.

The unit, exclusive of batteries, measures five inches by five inches by two inches.

CHAPTER I

INTRODUCTION

The research reported herein consisted of the development, construction and evaluation of an eight-channel, transistorized commutator for the Electrical Engineering School of the Georgia Institute of Technology. The motives underlying the school's need for the commutator were: to examine the feasibility of a thirty-two channel transistorized commutator for the school's speech analysis equipment, and to obtain a classroom teaching aid. The device allows up to eight independent voltage waves to be displayed as portions of an oscilloscope trace.

A number of transistorized commutators have been reported in the literature (1,2,3), but not in sufficient detail to allow immediate construction of a reliable working unit.

It was decided from the start that general purpose diodes and transistors would be used if possible, so that the cost per commutated channel would be kept low.

There are many circuits possible for a commutator, but they will fall loosely into two categories. Either they will be complex switches (diode or transistor trees, matrices, etc.) controlled by simple synchronizing circuits, or they will be simple switches (a single diode or transistor in each input channel, for instance) controlled by complex wave forming circuits. The commutator of this thesis is in the first category. It is a matrix of thirty-two point-contact diodes, controlled by a chain of three bistable multivibrators. The matrix is followed by a cascaded, two-transistor emitter follower, which provides a low output impedance for the commutated signal. Figure 1 shows in block diagram the functional arrangement of the parts of the commutator.

A fast-sampling pulse stretcher was developed to improve the output presentation at the highest commutation rate of 150,000 samples per second, but it is not included in the unit which was constructed.

In the following chapters, the completed unit is first described, then the matrix, multivibrators, emitter followers, and pulse stretchers are further discussed. The appendix includes a tabulation of the formulas (4) needed to design a non-saturating, transistorized, bistable multivibrator of the type used in the commutator.

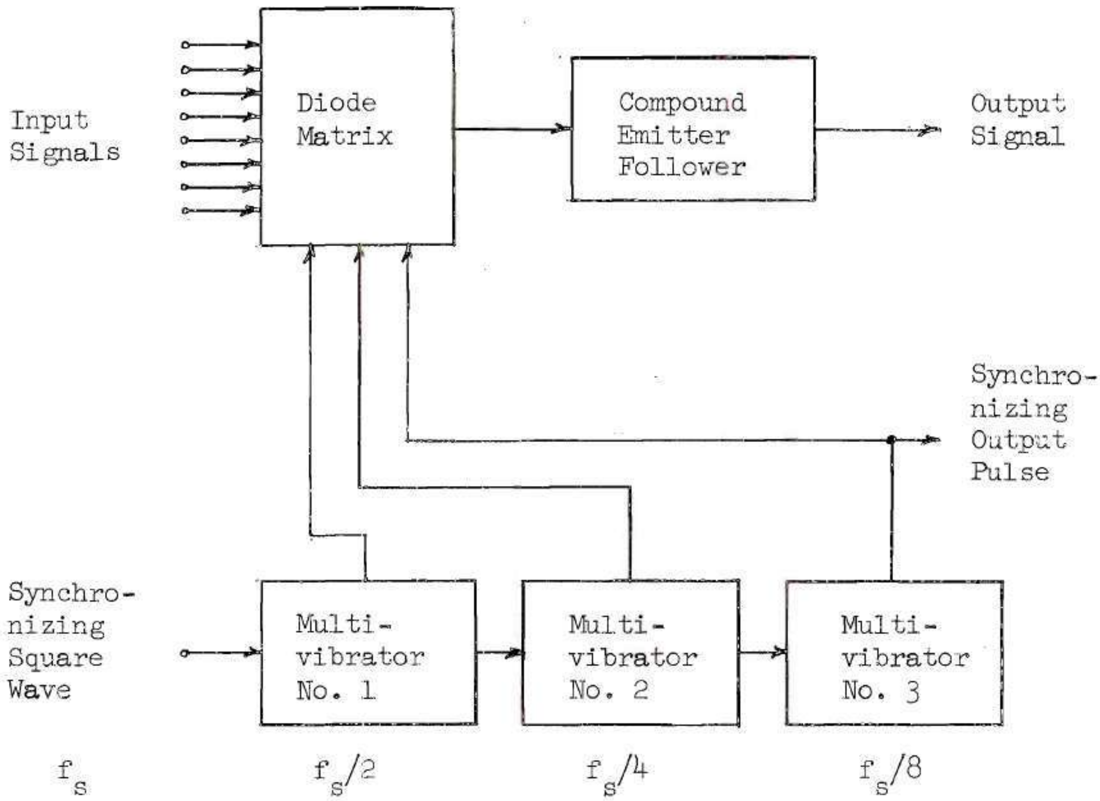


Figure 1. An Eight Channel Commutator.

CHAPTER II

THE EIGHT-CHANNEL COMMUTATOR

The schematic diagram of the eight-channel commutator constructed for the Electrical Engineering School is shown in Figure 2. The diode matrix is in the upper left, the control multivibrators are at the bottom, and the output emitter follower is at the upper right of the diagram. Figure 3 is a photograph of the completed unit.

The matrix, control circuits, and emitter follower were constructed on separate fiberglass boards so that improved versions of any of these three subdivisions might be substituted in the future. Commercial diode clips and universal transistor sockets were used. The components in the control circuits and the emitter follower were mounted by insertion of their leads through holes drilled in the fiberglass board. The components were on the top of the board, and their leads served as binding posts for the wiring underneath.

The performance of the eight-channel commutator may be judged from the oscilloscope waveforms of Figures 4 and 5, and from the linearity plot of Figure 6.

The output waveforms in Figure 4 show the effects of different input signal conditions. For the left-hand column, only d.c. input voltages were being sampled. In the right-hand column are shown the output waveforms when a combination of d.c. and sine wave inputs with a wide range of frequencies were commutated.

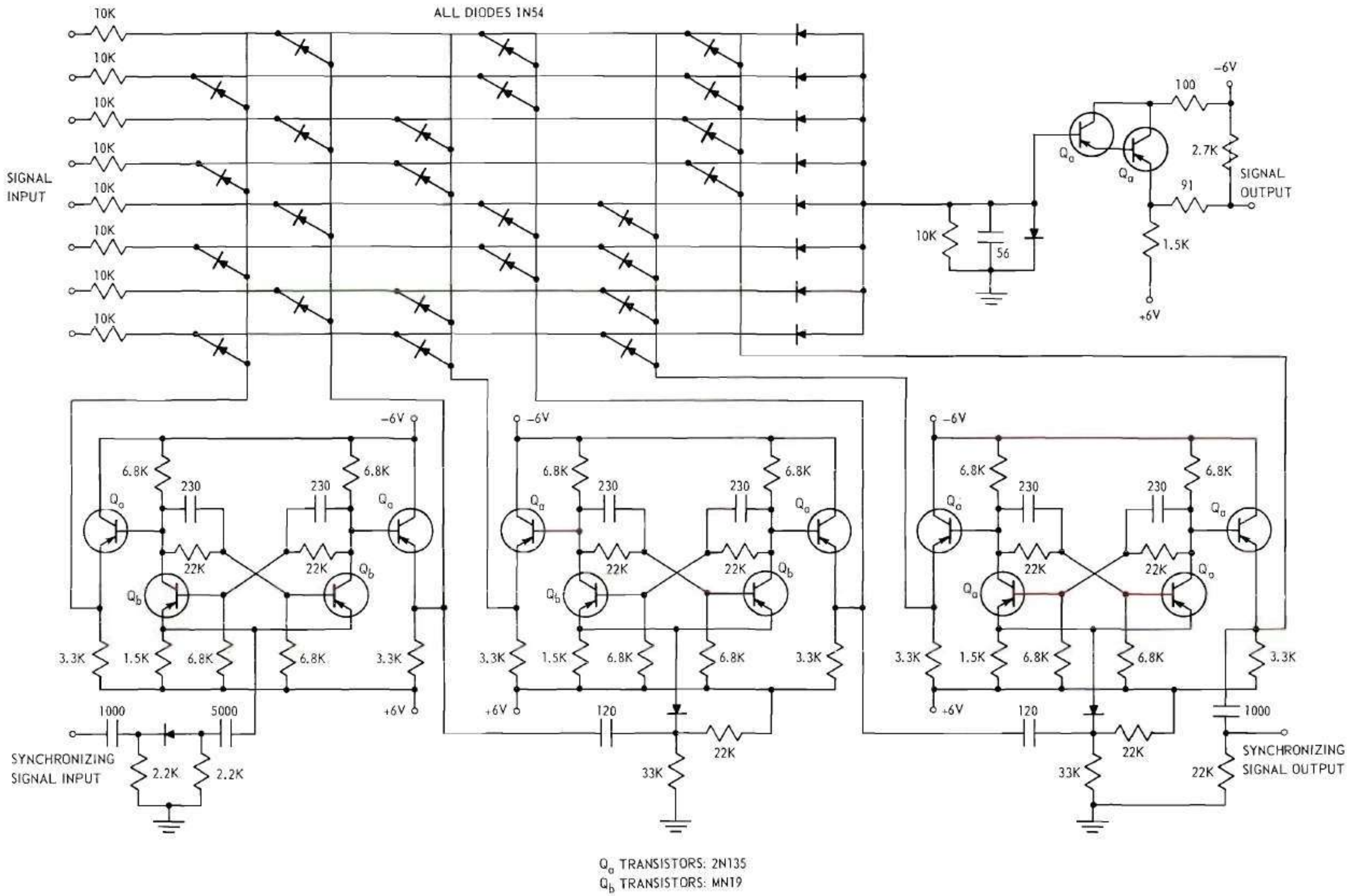
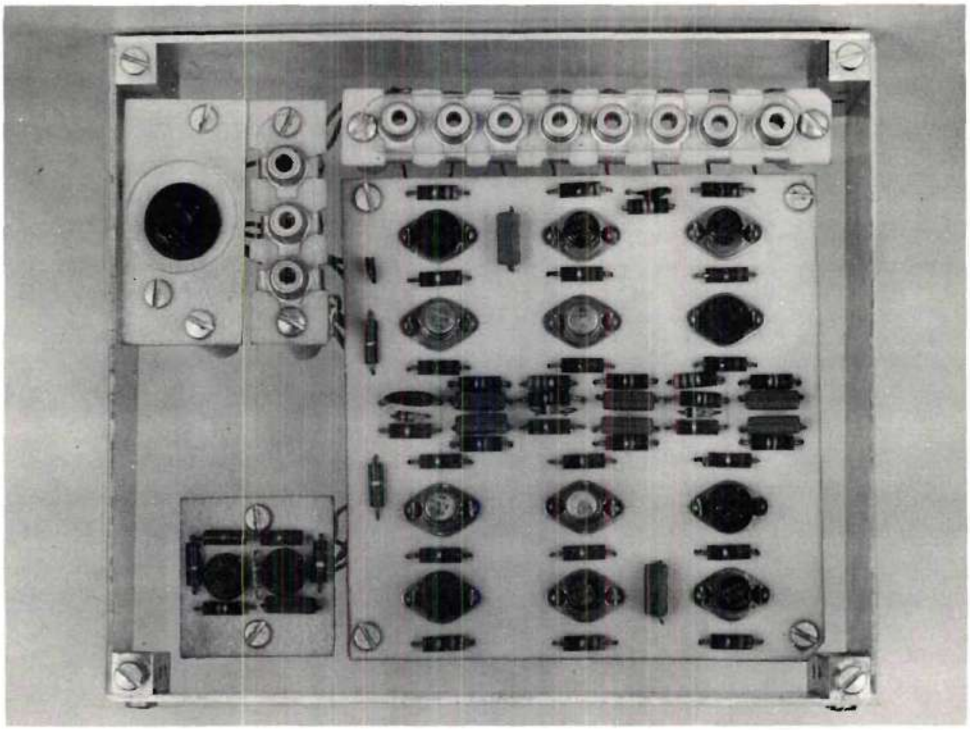
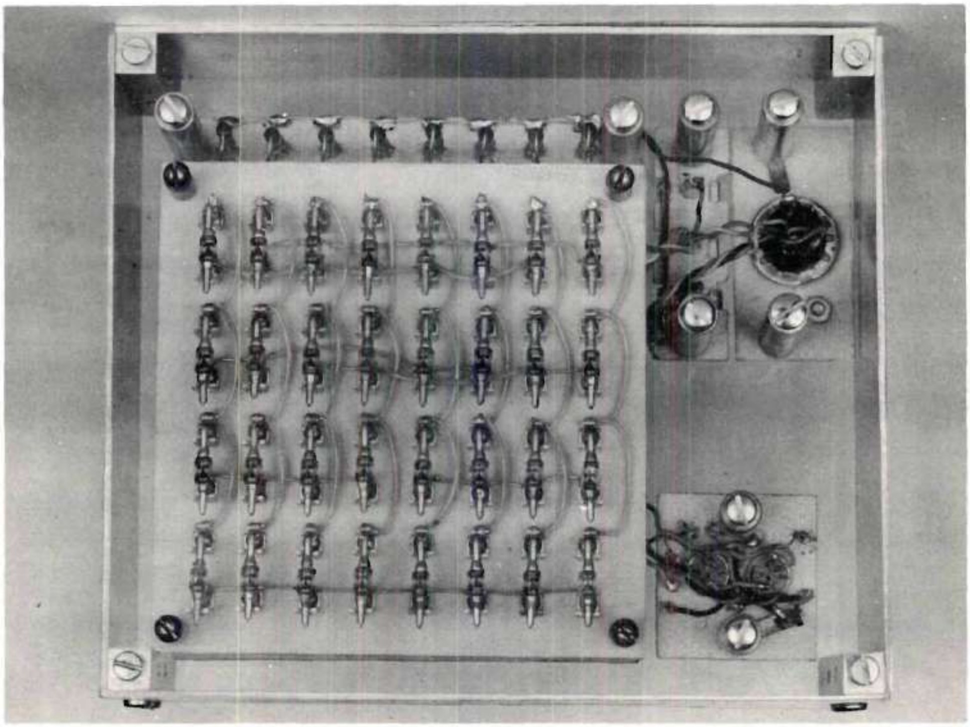


Figure 2. Schematic Diagram of an Eight Channel Commutator.

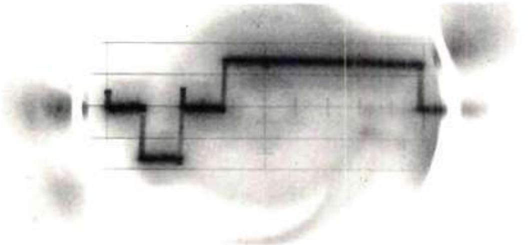


(a) Top View

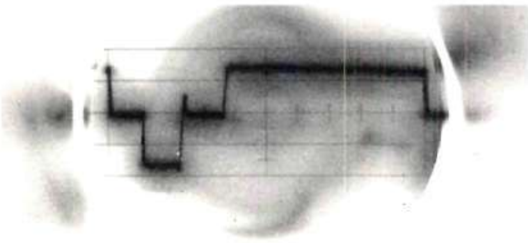


(b) Bottom View

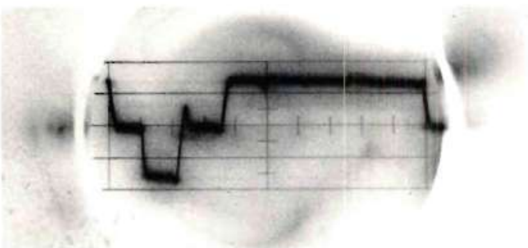
Figure 3. The Commutator Unit.



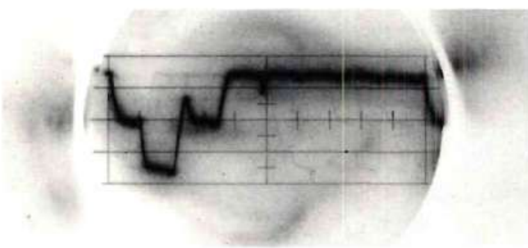
(a) 1.5, 3.0 and 1.5 Volts dc,
10 KC Sampling.



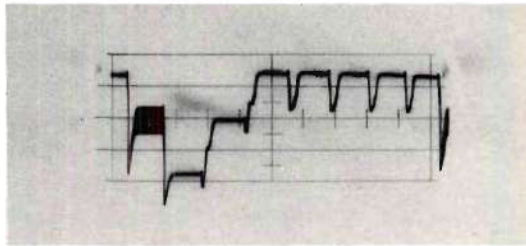
(b) 1.5, 3.0 and 1.5 Volts dc,
50 KC Sampling



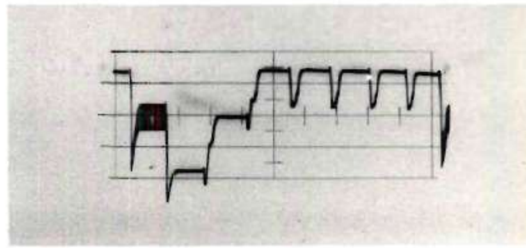
(c) 1.5, 3.0 and 1.5 Volts dc,
100 KC Sampling.



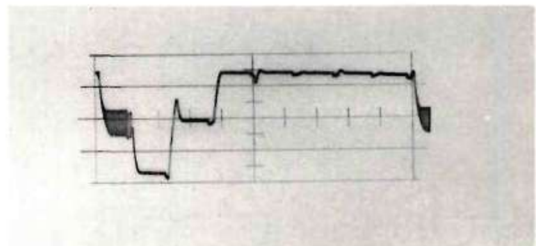
(d) 1.5, 3.0 and 1.5 Volts dc,
150 KC Sampling.



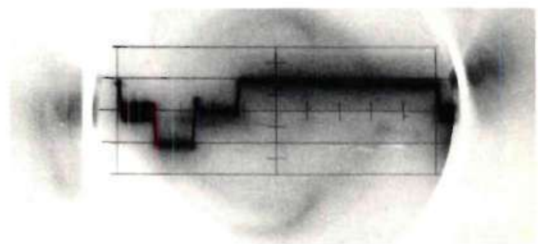
(e) 0.3 Volt rms at 100 cps, 1.5 Volts
Bias, 100 KC Sampling.



(f) 0.3 Volt rms at 10 KC, 1.5 Volts,
Bias, 100 KC Sampling.

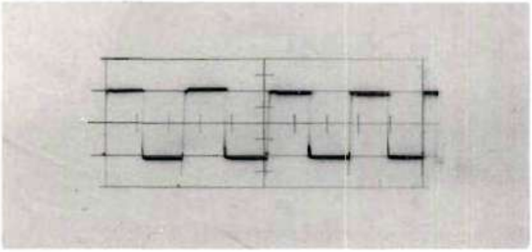


(g) 0.3 Volt rms at 50 KC, 1.5 Volts
Bias, 100 KC Sampling

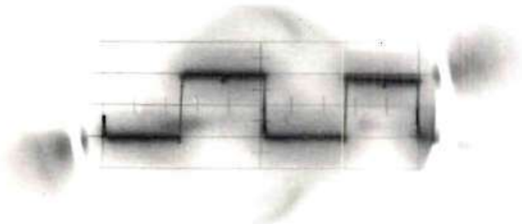


(h) 0.3 Volt rms at 5 KC, 1.5 Volts Bias, also
3.0 and 1.5 Volts dc, 50 KC Sampling.

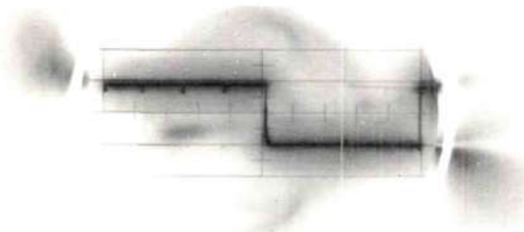
Figure 4. Commutator Output Waveforms.



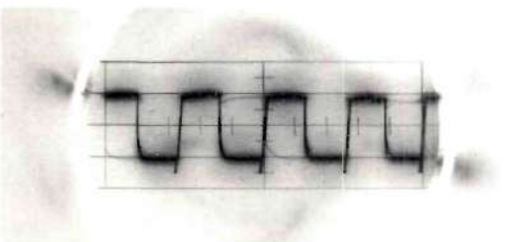
(a) Multivibrator No. 1, 10 KC.



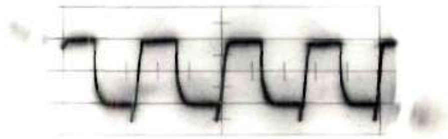
(b) Multivibrator No. 2, 10 KC.



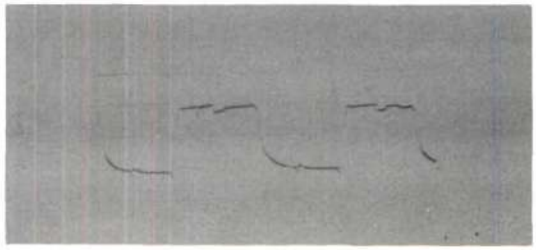
(c) Multivibrator No. 3, 10 KC.



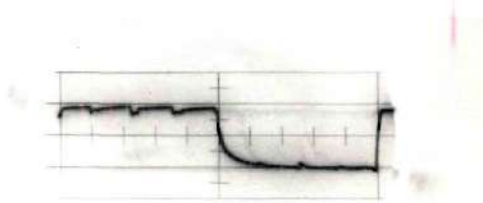
(d) Multivibrator No. 1, 50 KC.



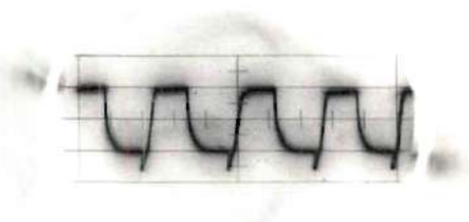
(e) Multivibrator No. 1, 100 KC.



(f) Multivibrator No. 2, 100 KC.



(g) Multivibrator No. 3, 100 KC.



(h) Multivibrator No. 1, 150 KC.

Figure 5. Multivibrator Waveforms.

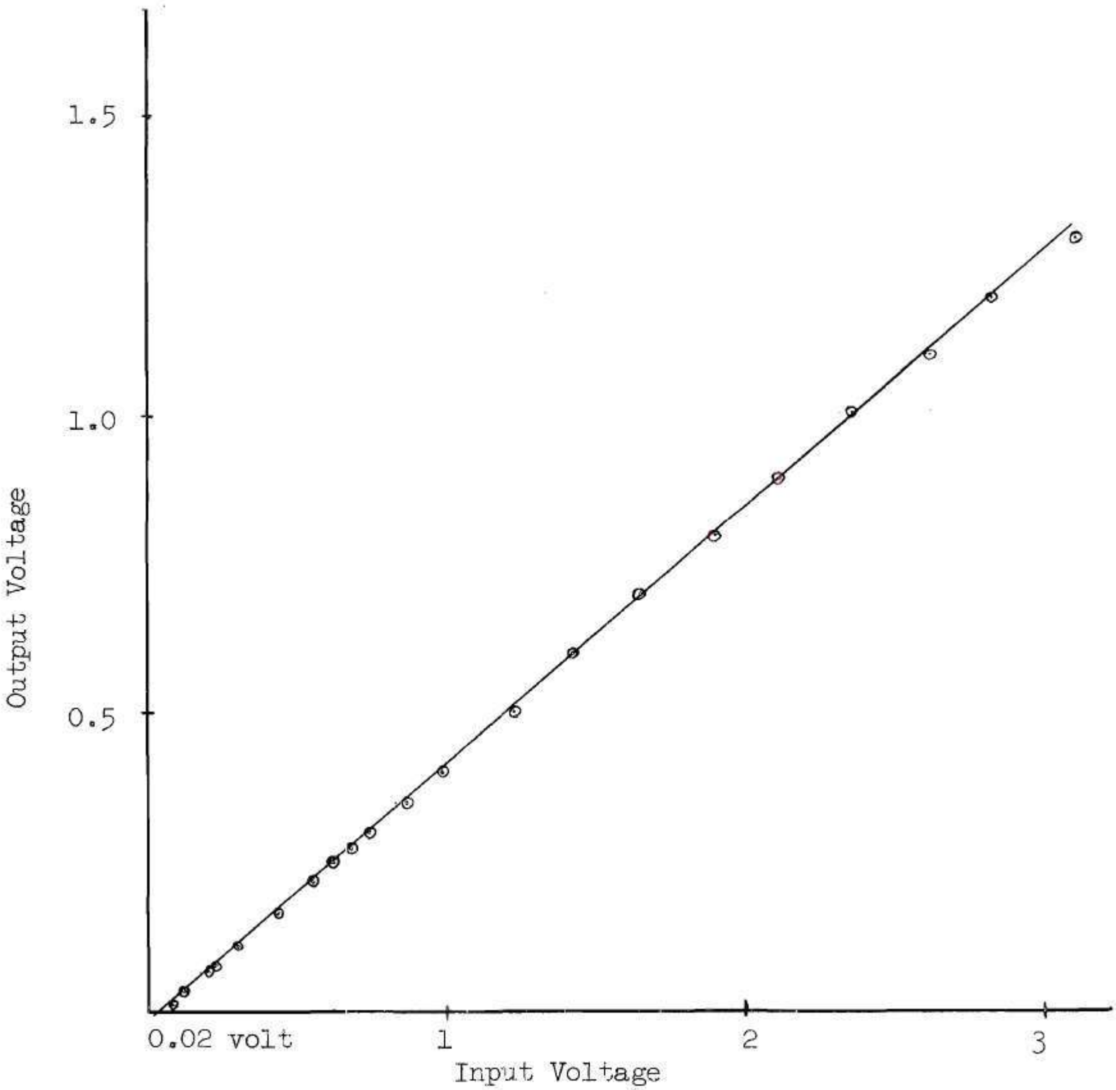


Figure 6. Output Voltage _versus_ Input Voltage.

The waveforms from the control multivibrators are shown in Figure 5. Rise and fall times are 1.0 microsecond. The peak-to-peak voltage swing is 8.0 volts.

The lower limit of linear operation of the commutator, denoted as 0.02 volt in Figure 6, is the extent of disparity in the output signal samples when the inputs are all short circuited. The upper limit of 3.0 volts is the point of 1.0 per cent departure from linear transmission.

The diodes of the commutator matrix, and the diode clamp at the input to the compound emitter follower, are oriented for negative input signals only. This arrangement was chosen to facilitate development of the pulse stretcher described in Chapter 6. Reversal of the diodes would permit commutation of positive signals rather than negative signals.

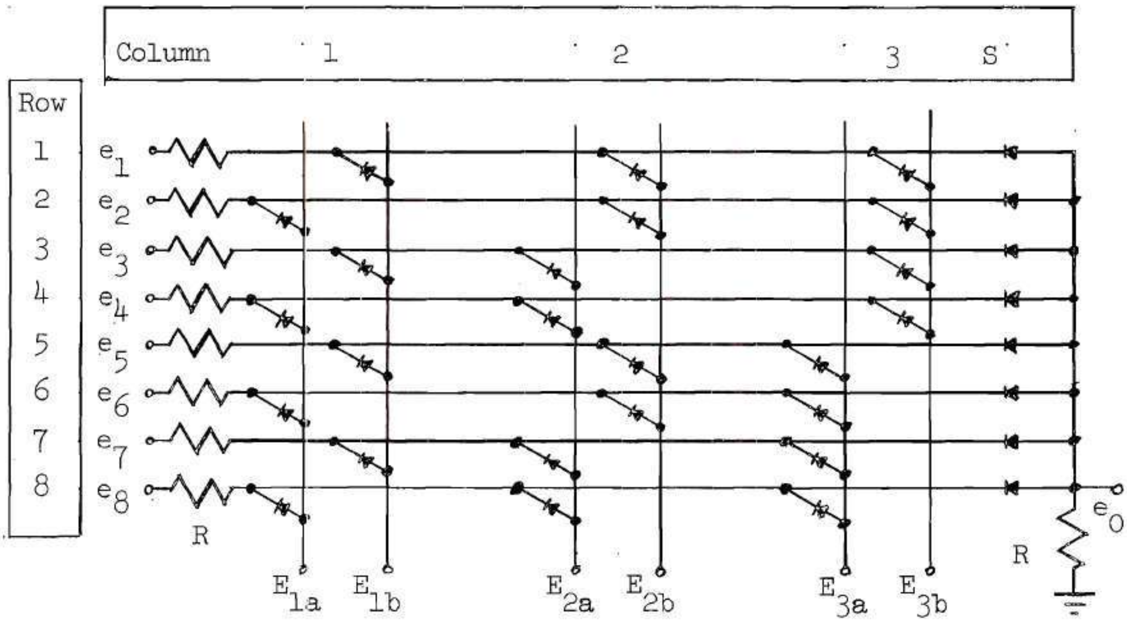
CHAPTER III

THE DIODE MATRIX

The matrix of thirty-two point-contact diodes which constitutes the commutator switch was analyzed in 1949 by Brown and Rochester (5). The matrix and the schedule of control voltages needed for commutation are shown in Figure 7.

The switching operation of the matrix can be traced out with the aid of the voltage schedule. Suppose, for instance, the control voltages are as in the first row of the schedule. The diodes connected to the positive voltages, E_{1a}, E_{2a}, and E_{3a}, are all forward biased, if the control voltage is more positive than the signal voltage in each case. The input signals, e_2, e_4, e_6, e_8, e_3, e_7, and e_5 are attenuated by the shunting action of the source resistances of the control voltage generators. In addition, the diodes in column S which are in rows 2 through 8 are back-biased by the positive control voltages, and these diodes also isolate the signals e_2 through e_8 from the output. Only e_1 can be connected to the output, and that will occur only if e_1 is negative, in which case the first-row diode in column S will be forward biased.

All of the negative control voltages, E_{nb}, serve to back-bias the diodes to which they connect, if the control voltages are more negative than half the voltage of each input signal. Similar analyses for the other voltage conditions of the schedule indicate the way the matrix commutates.



(a) Diode Matrix Switch

E_1		E_2		E_3		e_0
a	b	a	b	a	b	
+	-	+	-	+	-	e_1
-	+	+	-	+	-	e_2
+	-	-	+	+	-	e_3
-	+	-	+	+	-	e_4
+	-	+	-	-	+	e_5
-	+	+	-	-	+	e_6
+	-	-	+	-	+	e_7
-	+	-	+	-	+	e_8

(b) Control Voltage Schedule

Figure 7. Eight-Channel Diode Matrix Switch.

The matrix of Figure 7 will switch only negative input signals, and it is necessary that the magnitude of the negative control voltages exceed one-half the magnitude of the input signals. In addition, the positive control voltages must be positive with respect to ground.

Positive going signals could be commutated by the matrix if every diode were reversed.

The matrix configuration is adaptable to commutators with 2^n input channels (these are the rows in Figure 7). The binary symmetry of the matrix is such that there are $\underline{n}$ balanced-to-ground control circuits (the numbered columns in Figure 7). There are $(n+1)(2^n)$ diodes in each matrix. The voltage schedule in Figure 7 indicates that the frequency of the $\underline{n}^{th}$ control-circuit voltage is half the frequency of $(n-1)^{th}$ control-circuit voltage.

The matrix for a 16-channel switch would have sixteen rows instead of eight, and five columns instead of four. The sixteen diodes in column No. 1 of the larger commutator would alternate between the balanced halves of the control circuit; the diodes in the second column would alternate by pairs; and the diodes of the third column would alternate by fours, as in Figure 7. The diodes in the fourth column, the added balanced-to-ground control circuit, would alternate between the two control lines by eights. The fifth column would be sixteen series diodes, one in each signal path, as in Figure 7.

Still larger matrices, with thirty-two channels, sixty-four channels, etc., would follow the pattern of symmetry established above.

The diodes in the first $\underline{n}$ columns of the matrix serve either to shunt the (n-1) undesired signals or to disconnect the control voltages from the circuit of the desired signal. The n series diodes (the column marked S in Figure 7) serve to disconnect the output from all except the desired channel.

Figure 8 is an equivalent circuit which illustrates the on and off conditions for a particular input channel. In the actual circuit, the values of forward diode resistance, R_{f1} and R_{f2}, will be different, and so will the values of back resistance, R_{b1} and R_{b2}. The actual resistance will depend on how many diodes, biased in the same direction, are in parallel in the actual circuit. The output impedance of the controlling circuit, which is in series with the forward conducting diodes, is also involved in the actual value of R_f.

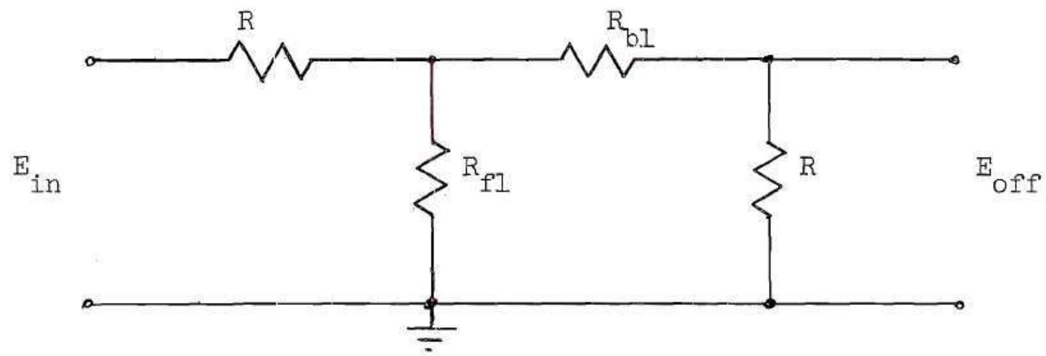
In spite of the limitations of the equivalent circuits of Figure 8, it is of interest to derive the signal leakage ratio E_{off}/E_{on}, assuming that $R_{f1} = R_{f2}$, $R_{b1} = R_{b2}$, and that the control circuit impedance is zero. Then,

$$\frac{E_{off}}{E_{on}} = \frac{[\ R_t(R + R_b) - R_b^{\ 2}\ ]\ R_f}{[\ R_t(R + R_f) - R_f^{\ 2}\ ]\ R_b}, \qquad (1)$$

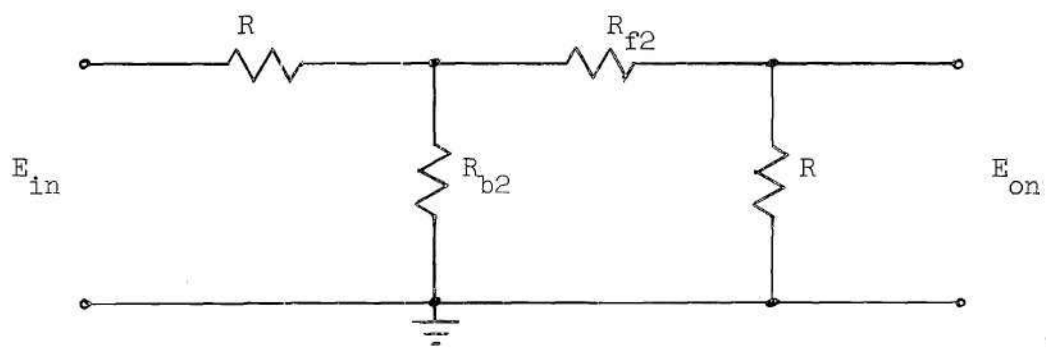
where

$$R_t = R_b + R + R_f \,.$$

Equation 1 can be easily evaluated for each of three conditions imposed on R; namely, either that $R = R_b$, or $R = R_f$, or $R = \sqrt{R_b R_f}$. It may be assumed that $R_b \gg R_f$, and even that $\sqrt{R_b R_f} \gg R_f$. Now when



(a) Off Condition



(b) On Condition

Figure 8. Equivalent Circuit for a Matrix Channel.

$R = R_b$, or when $R = R_f$,

$$\frac{E_{off}}{E_{on}} \cong \frac{3R_f}{2R_b} , \tag{2}$$

and when $R = \sqrt{R_b R_f}$,

$$\frac{E_{off}}{E_{on}} \cong \frac{2R_f}{R_b} . \tag{3}$$

Equations (2) and (3) indicate that the value of R is not critical, so long as it lies between the values of R_f and R_b, and so long as $R_b \gg R_f$.

The resistances of the 1N54A diodes used in the matrix were measured with an ohmmeter. The static values of R_b (at 7 volts back bias) were between 2 and 10 megohms. The values of R_f (at 12 milliamperes forward current) were between 40 and 100 ohms. Measurements were then made on one of the diodes at conditions more closely approximating the matrix operation. The static value of R_b at 1 volt back bias was 1.5 megohms. R_f at 0.5 milliampere forward current was 300 ohms. The latter values indicate that a static signal leakage ratio (E_{off}/E_{on}) of approximately 4×10^{-4} might be expected. The leakage of signal during a switching transient would be much larger, of course.

Measurements were made to check the computed leakage ratio. A static control voltage schedule was maintained on the matrix with batteries, and a sine wave signal, biased at -1.5 volts, was introduced into the "on" channel. The amplitude of the test signal was adjusted to make the output, E_{on}, be 0.5 volt rms. Then the test signal was introduced into any "off" channel, and the output, E_{off}, was measured. It was 0.00016 volt. The signal leakage was thus 3.2×10^{-4} by measurement. This value held for signal frequencies from 10 to 100,000 cycles per second.

In addition to the leakage characteristic, the switching speed of the matrix is important. The time required for a single 1N54 point-contact diode to be switched from 0.5 milliamperes forward current to 1.0 volt back bias was found to be 0.1 microsecond. The switching time of the matrix of thirty-two diodes was next measured. Two control circuits were held at fixed voltages and the third control circuit was driven by a square wave with a 0.03 microsecond rise time. The matrix switching was surprisingly fast: 0.4 microsecond. This switching speed held for drive frequencies from 10,000 to 100,000 cycles per second.

In summary, the diode matrix leakage ratio was found to be 70 decibels for signals up to 100,000 cycles per second, and the switching time was 0.4 microsecond.

The fast switching capabilities of the diode matrix are not fully utilized in the eight-channel commutator. The switching times of the multivibrator control stages, which are described in the next chapter, constitute the performance limitation. However, a multivibrator circuit is described which had switching times as low as the diode matrix, but which required a larger number of transistors. The choice of the control circuit used in the commutator was based on its economy of parts rather than its speed.

CHAPTER IV

THE TRANSISTOR MULTIVIBRATOR CONTROL CIRCUIT

Introduction.--The control voltages for the diode matrix are obtained from a chain of three transistor bistable multivibrators. Each multivibrator except the first is driven by a preceding multivibrator; the first in the chain is driven by an external square-wave generator. Each multivibrator runs at half the frequency of its drive voltage.

In Chapter III it was pointed out that a sixteen-channel matrix would have four control circuit pairs, a thirty-two channel matrix would have five pairs, etc. Each additional control circuit would require one additional multivibrator in the chain. This is the primary advantage of this mode of commutation: that only n control stages are needed for 2^n commutated input channels. The disadvantage is that the switching times of the multivibrators contribute additively to the total switching time of the commutator.

In the eight-channel commutator, each multivibrator completes its switching approximately one microsecond after it is triggered. Examination of the voltage schedule in Figure 7 will show that the longest switching transient should be expected when the commutator is switching from channel No. 4 to No. 5, and from No. 8 to No. 1. This was verified experimentally. The transient time was about 2.5 microseconds.

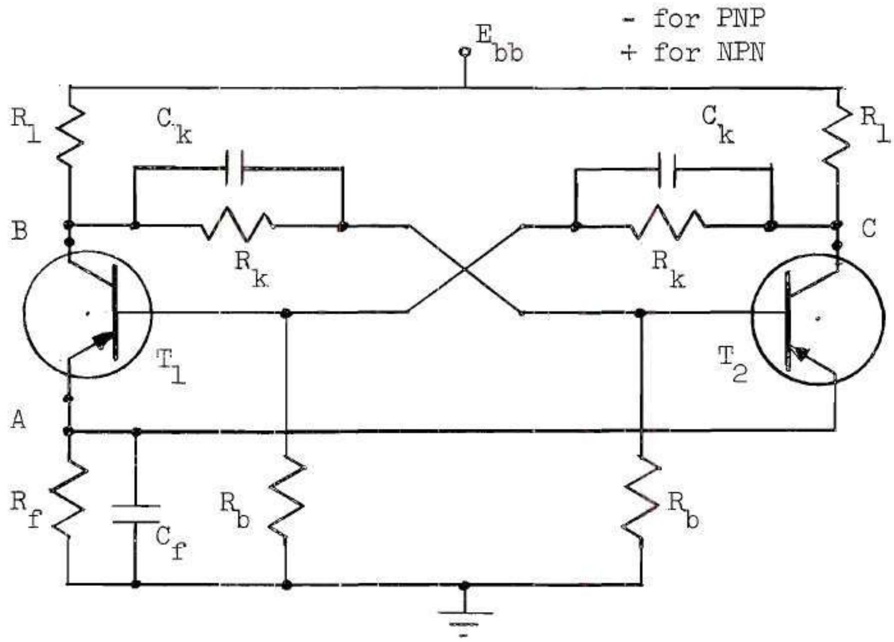
Static circuit design.--A useful multivibrator synthesis procedure, worked out by Suran and Reibert (4,6), was used to design a free-running multivibrator, a complementary symmetry bistable multivibrator, and a non-saturating bistable multivibrator. The latter circuit was used in the commutator.

Only a sketchy review of the synthesis will be given; the reader is referred to the literature for details.

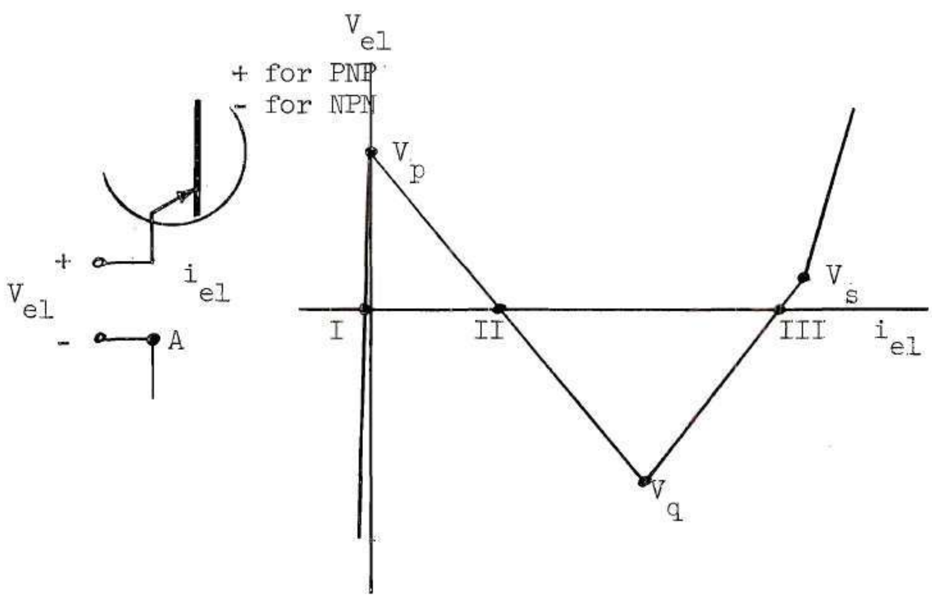
Figure 9(a) is a basic transistor multivibrator circuit. When one transistor is conducting, it causes the other transistor to be held off, but if either transistor is caused by an injected signal to change its state, regeneration speeds the process and the transistors rapidly exchange the roles of on-off conditions.

The circuit may be analyzed by breaking it at point A in Figure 9(a) and finding the static volt-ampere characteristic at the two terminals which result. Figure 9(b) is a typical curve. The region of reverse current represents cut-off for transistor T_1; the negative-slope portion of the curve represents regeneration, when both transistors are characterized by gain; the point V_q marks cut-off of T_2; and V_s is the saturation point of T_1.

Closing the circuit at A is equivalent to making the current (i_e) axis the load line on the characteristic. The circuit is stable at two points, I and III. It is generally desirable to have the conduction stop short of steady-state saturation to avoid carrier storage effects, which increase switching time. From Figure 9 it is seen that if V_s is opposite in sign to E_{bb}, saturation is avoided.



(a) Schematic Diagram



(b) Static Characteristic

Figure 9. A Bistable Multivibrator.

The voltage-current values for the break points of the curve were determined by Suran and Reibert from the relationships of the resistances and the static currents and voltages in the circuit. The synthesis involves seven unknowns, E_{bb}, R_1, R_b, R_k, R_f, C_k, and C_f. Of these, however, E_{bb} at least will usually be specified. In addition the collector voltage swing, the maximum collector current, and the circuit recovery time are convenient parameters which may be specified. The formulas developed by Suran and Reibert for a bistable circuit are presented in the Appendix, with the computation of the multivibrator stages of Figure 2 as an example.

<u>Rise time design problems</u>.--Optimum rise and fall times are desirable in the multivibrator. An approximation of the theoretical maximum pulse repetition rate (6) is $2f_\alpha / \sqrt{\beta_o}$, and the corresponding rise time is about one-fourth the period, or

$$\tau_{rise} \cong \frac{\sqrt{\beta_o}}{8f_\alpha} \tag{4}$$

where

β_o = d.c. current gain, and

f_α = alpha cutoff frequency.

For example, the MN19 transistor used in the commutator, which has a β_o of 40 and f_α of 8.0 megacycles, should have a rise time of about 0.1 microsecond. This can serve only as a first order estimate, however. Experimental rise times will be longer.

The circuit in Figure 9 may be triggered at point A, or at points B and C. When frequency halving is desired, triggering at point A is convenient. Frequency halving is also possible with triggers connected to B and C, but steering diodes are necessary in this case.

It has been shown (6) that the capacitor, C_f, serves to speed up the multivibrator transition by bypassing the common emitter resistance. If it is not bypassed, the emitter resistance decreases the loop gain by degeneration. When the circuit is triggered at point A, however, C_f must be removed, and it is then necessary that the triggering circuit source impedance be low, to replace the bypass function of C_f. The emitter follower which couples the multivibrator to the diode matrix, as shown in Figure 2, serves also as a convenient low-impedance source of trigger voltage to drive the succeeding multivibrator.

The values of the coupling capacitors, C_k, were found by trial-and-error. The observed waveform of a multivibrator in the completed commutator is sketched in Figure 10. The rising part of the wave corresponds to transistor turn-on, and it was possible to steepen it by increasing the value of C_k. The falling edge of the wave consists of two parts: a very steep initial fall corresponding to regenerative action when both transistors are active, and a slower, exponential fall corresponding to the charging of C_k through the load resistance, R_l, after one of the transistors cuts off. Increasing C_k increased the charging time, as shown in Figure 10. The value of C_k was chosen so as to balance the improvement desired in the turn-on time against the degradation suffered in the charging time of C_k. With 250 micromicrofarad coupling

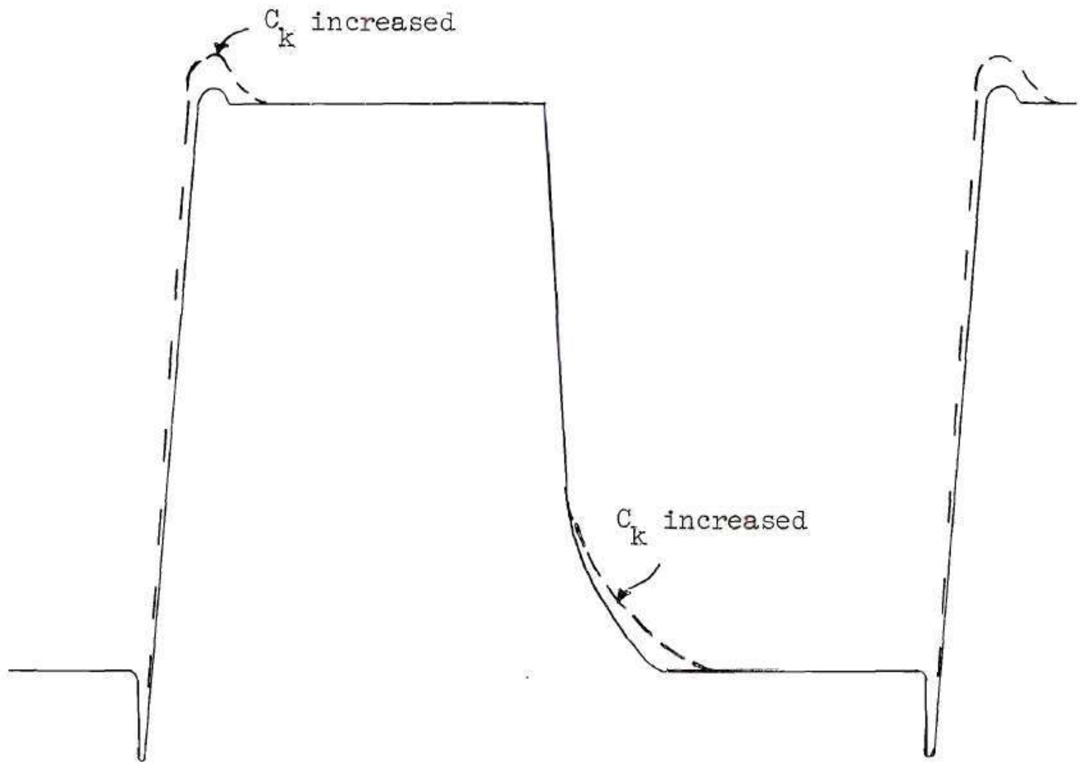
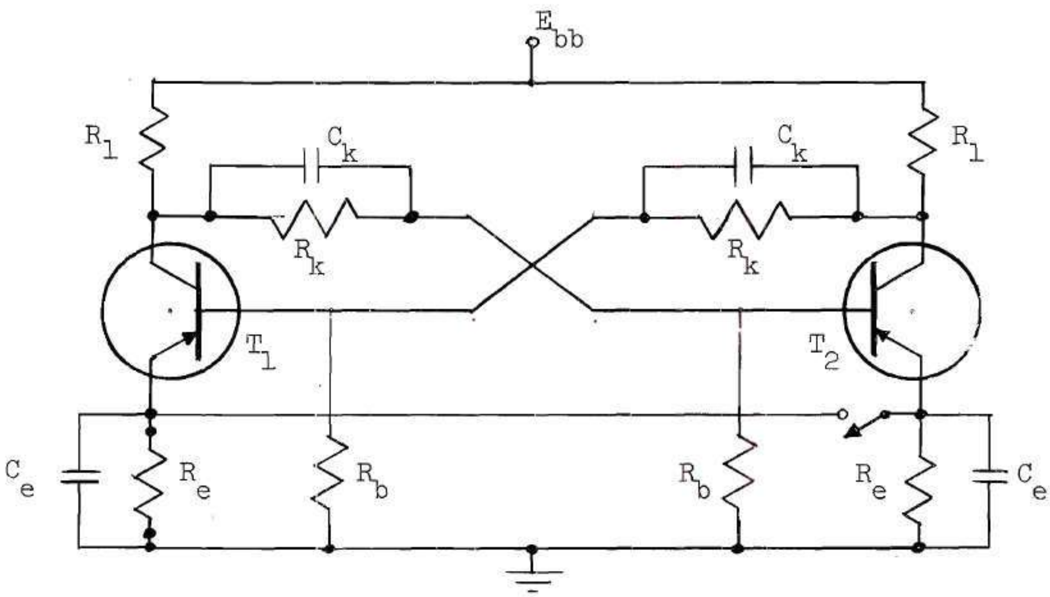


Figure 10. Collector Voltage in a Bistable Multivibrator.

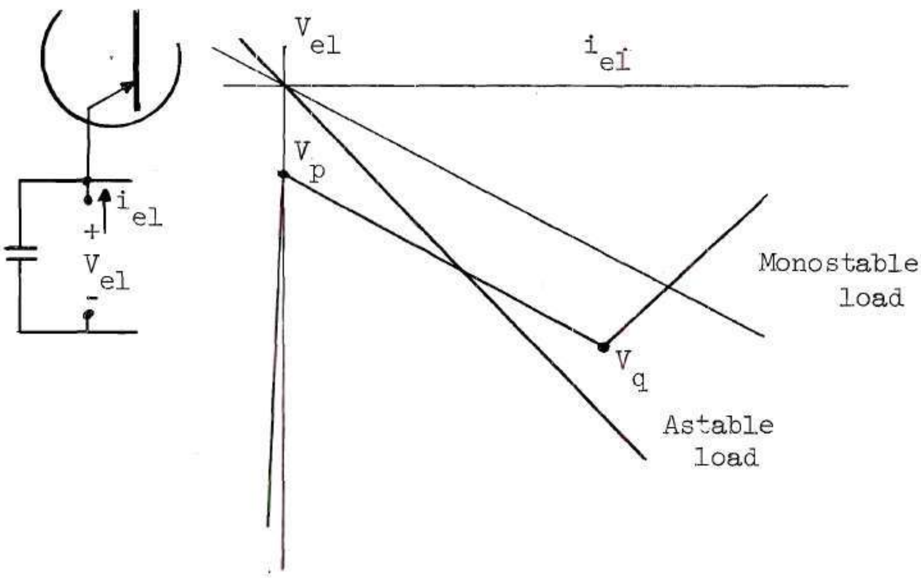
capacitors, turn-on time is 1.0 microsecond. The load resistors are 6800 ohms, and the charging time constant, R_1C_k, is thus 1.7 microseconds. The steep portion of the falling characteristic represents a time of less than 0.5 microsecond. Photographs of the multivibrator waveforms in the completed commutator are reproduced in Figure 5.

<u>Other multivibrator circuits</u>.--If the emitters in the circuit in Figure 9 are returned to ground through separate, equal resistors, as in Figure 11(a), the multivibrator will be either free-running or monostable. The analyses of the free-running and the monostable circuits are also treated by Suran and Reibert (4). Removal of one of the emitter resistances permits determination of the circuit's volt-ampere characteristic, which is found to be like that shown on Figure 11(b), and reinsertion of the resistor corresponds to a load line that, depending on its magnitude, intersects either the negative resistance region or the positive resistance region of the characteristic. Intersection with the negative slope portion of the curve determines a free-running circuit; intersection with the positive slope portion effects monostable operation. A circuit was constructed which permitted, by the switch arrangement shown in Figure 11(a), either bistable or free-running operation. If this arrangement were used as the first control stage in the commutator, the square wave signal generator could be optional.

The bistable multivibrator shown in Figure 12 has complementary symmetry (6). It is characterized by high efficiency (the ratio of power delivered to battery power drain). This circuit was constructed, and



(a) Schematic Diagram



(b) Static Characteristic

Figure 11. Monostable or Astable Multivibrator.

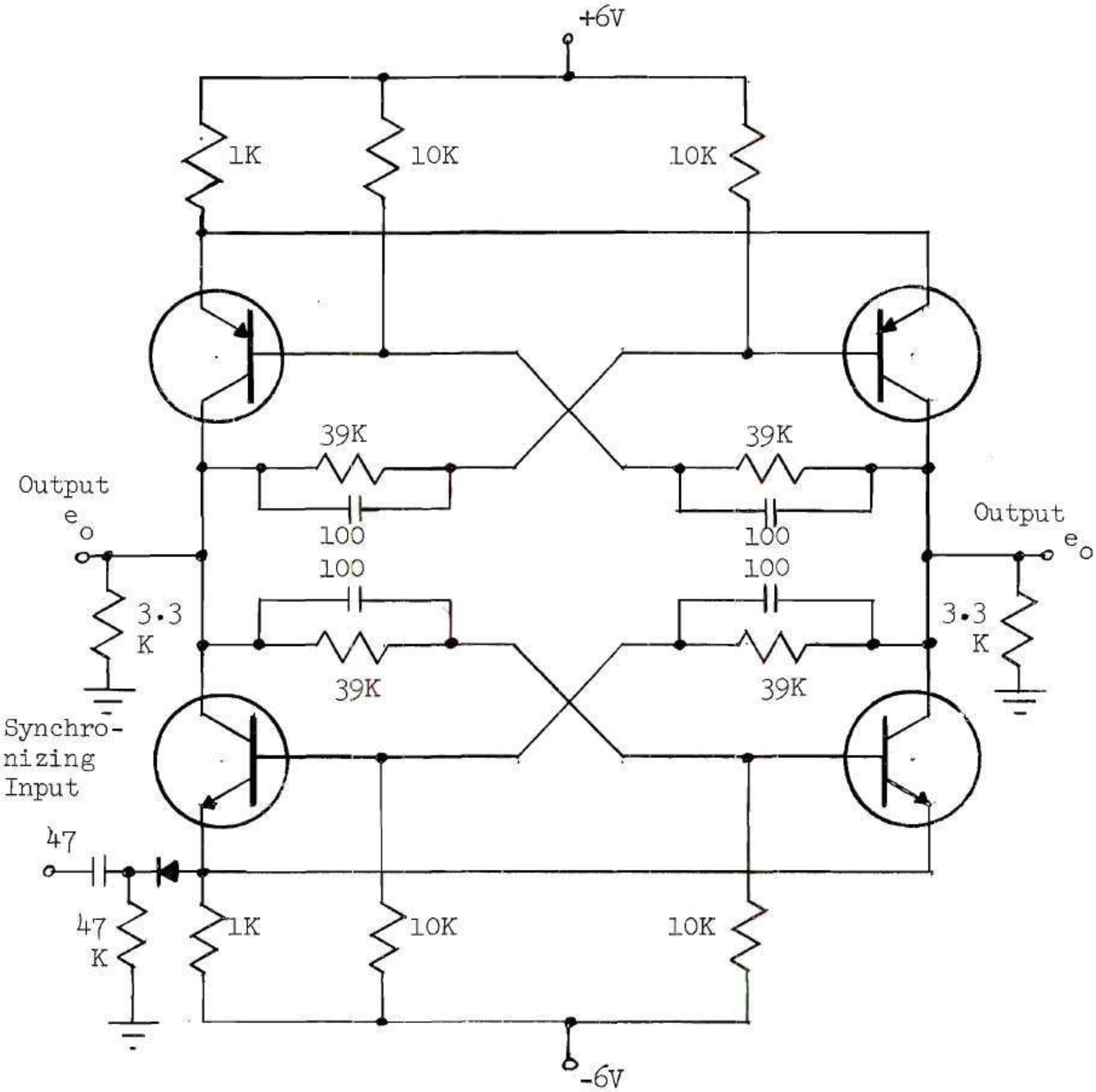


Figure 12. A Complementary Symmetry Bistable Multivibrator.

rise and fall times of 0.35 microsecond were measured. Frequency halving was accomplished by introducing the trigger voltage at point A in Figure 12. The complementary-symmetry circuit, of course, required four transistors, twice as many as the basic circuit of Figure 9.

CHAPTER V

THE EMITTER FOLLOWERS

Direct coupled emitter followers connect the multivibrators to the diode matrix, and a compound, or Darlington circuit (6), emitter follower is used as an output stage for the commutator. The emitter followers between the matrix and the control circuits are needed to prevent the input signals from affecting the multivibrator operation. The output emitter follower provides a low output impedance.

Figure 13(a) is an equivalent diagram of an emitter follower driven at its base by a voltage generator, v_g, which has a source impedance, r_g. The low frequency output impedance of the emitter follower is

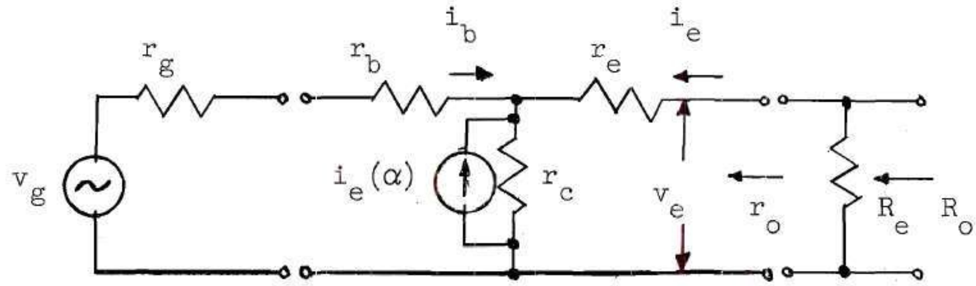
$$R_o = \frac{R_e r_o}{R_e + r_o} , \tag{5}$$

but

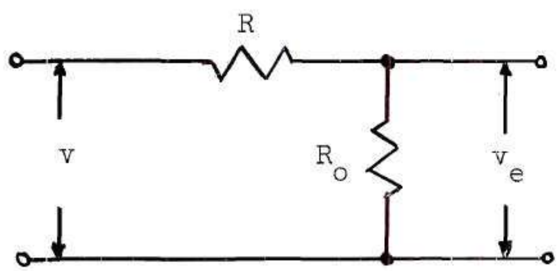
$$r_o = \frac{V_e \text{ (open circuit)}}{i_e \text{ (short circuit)}} . \tag{6}$$

Now

$$v_e\text{(open circuit)} = v_g r_c/(r_b + r_c + r_g) , \tag{7}$$



(a) General Emitter Follower



(b) Effect of Input Signal

Figure 13. Emitter Follower Equivalent Circuit.

and

$$i_e(\text{short circuit}) = \frac{v_g r_c}{r_e(r_b + r_c + r_g) + r_c(1 - \alpha)(r_b + r_g)} , \quad (8)$$

so that

$$r_o = \frac{r_e(r_b + r_c + r_g) + r_c(1 - \alpha)(r_b + r_g)}{r_g + r_c + r_b} . \quad (9)$$

Usually $r_c \gg r_b \gg r_e$, and r_o can be approximated as

$$r_o \cong \frac{r_e + r_c(1 - \alpha)(r_b + r_g)}{r_c + r_g} , \quad (10)$$

and when $r_c \gg r_g$,

$$r_o \cong r_e + (1 - \alpha)(r_b + r_g) . \quad (11)$$

Equation 11 indicates that the coupling between the emitter circuit and the base circuit is reduced by the presence of the transistor, and one factor of the reduction of coupling is $(1 - \alpha)$. In the commutator, this isolation is further increased by the voltage divider action of the emitter resistor in series with the input resistor. Figure 13(b) is an equivalent circuit which shows the effect of an input signal voltage, $\underline{v}$, on the voltage at the emitter of the emitter follower which couples the multivibrator and the matrix. It can be seen from Figure 13(b) that

$$v_e = v \frac{R_o}{R + R_o} . \tag{12}$$

In the commutator, the transistor and circuit parameters can be approximated as:

$$r_e = 10 \text{ ohms,}$$

$$r_b = 200 \text{ ohms,}$$

$$\alpha = 0.95,$$

$$r_g = 5000 \text{ ohms, and}$$

$$r_c = 10^6 \text{ ohms.}$$

With these values substituted in Equation 10,

$$r_o = 10 + (0.05)(200 + 5000)(10^6)/(5000 + 10^6)$$

$$= 270 \text{ ohms.}$$

In Figure 2, R_e = 3300 ohms, so Equation 5 can be solved for R_o:

$$R_o = (270)(3300)/(270 + 3300)$$

$$= 250 \text{ ohms,}$$

and since R = 10,000 ohms,

$$v_e = v(250)/(10{,}000 + 250)$$

$$= v/(41).$$

Since the base-to-emitter junction of the emitter follower transistor is forward biased, the base voltage is approximately equal to the emitter voltage. The effect of the input voltage, $\underline{v}$, on the output of the multivibrator is thus reduced by the factor, 1/41. Had the diode matrix been driven directly from the collectors of the transistors in the multivibrators, the isolation would have been $r_g/(R + r_g)$ or only 2/3, and the input signal could easily have stopped the operation of the multivibrators.

The output impedance of the compound (Darlington) emitter follower can be computed from Equation 11. Let it be assumed that the two transistors in the compound circuit (refer to Figure 2) are identical. Then

$$r_o = r_e + (1 - \alpha)^2(r_b + r_g) + (1 - \alpha)r_b \ . \qquad (13)$$

The coupling between the emitter circuit and the input circuit is reduced, by virtue of the compound connection, by $(1 - \alpha)^2$. In the commutator r_g is again 5000 ohms, the parallel resistance of the input and output resistors in the matrix. Using the same values as before for the other parameters,

$$r_o = 10 + (0.05)(200) + (0.05)^2(200 + 5000)$$

$$= 33 \text{ ohms.}$$

In the circuit in Figure 2, R_e = 1500 ohms, and the impedance at the emitter is therefore

$$R_o = (33)(1500)/(33 + 1500)$$

$$= 32 \text{ ohms.}$$

A voltage divider connects the emitter to the negative supply, so that the output voltage will be zero when the input signal is zero. The values of the resistances in the divider are shown in Figure 2. The output impedance of the commutator can now be computed:

$$R_{oc} = (1500)(32 + 45)/(1500 + 32 + 45)$$

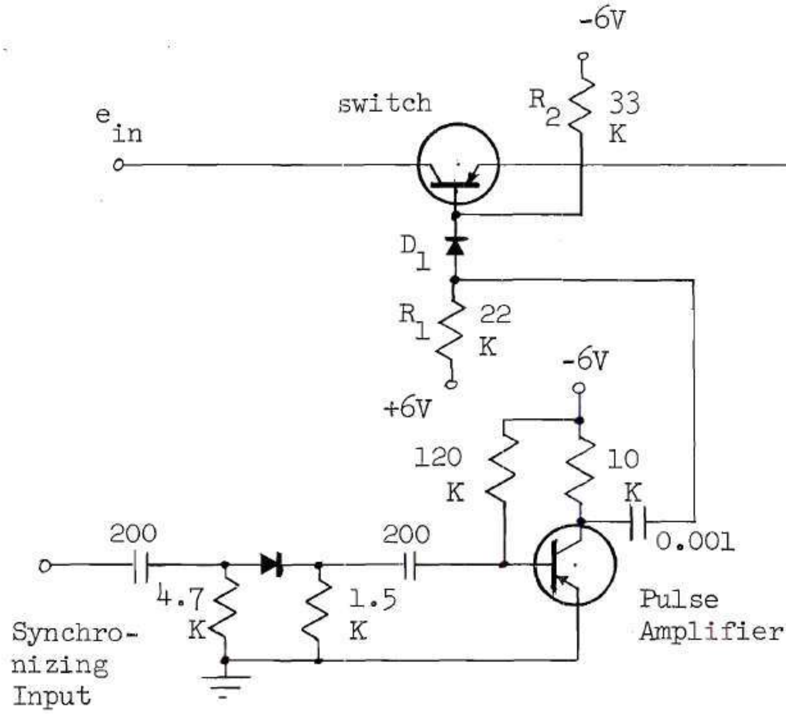
$$= 73 \text{ ohms.}$$

CHAPTER VI

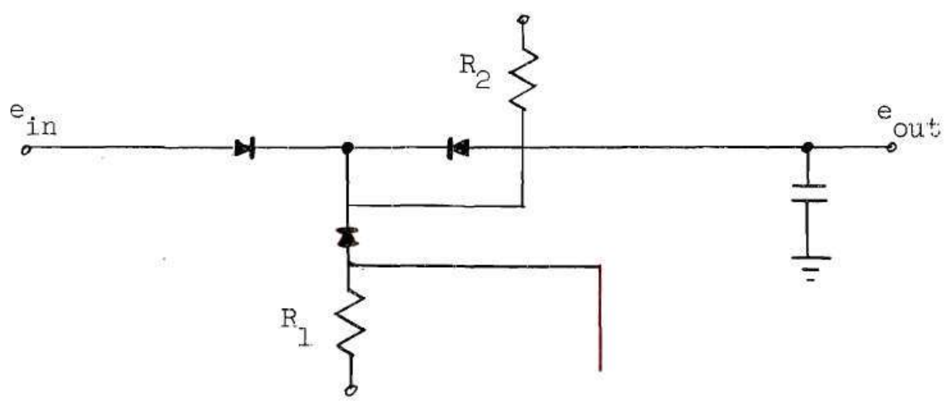
THE FAST-SAMPLING PULSE STRETCHER

At fast sampling rates the switching transient may last for an appreciable part of the time of a channel sample. The resulting presentation is ambiguous; one is not sure what part of the oscilloscope trace represents each input signal. However, if the "clear" part of the trace occurs always at the same place in each sampling cycle, it may be possible to resample the commutator output at that instant with a fast-sampling switch and to store the sample on a capacitor until the next "clear" output. This technique is sometimes called pulse stretching, and has been quite successful in removing the effects of contact bounce in mechanical commutators (7). In the pulse stretcher the storage capacitor is connected through the fast-acting switch to the commutator output for an instant of time just long enough to allow the capacitor to charge to the value of signal voltage from the commutator. It is clear that the output impedance of the commutator must be low, to allow rapid charging of the storage capacitor. Conversely, the impedance of the oscilloscope probe connected to the storage capacitor must be high, to avoid leakage of the stored charge between samples.

Figure 14(a) is a schematic diagram of the transistor pulse stretcher developed to be used with the eight-channel commutator. It was not included in the final model because it only improved the appearance of the output signal when sampling frequencies were above 100,000



(a) Schematic Diagram



(b) Transistor Switch Equivalent Circuit

Figure 14. A Fast-sampling Pulse Stretcher.

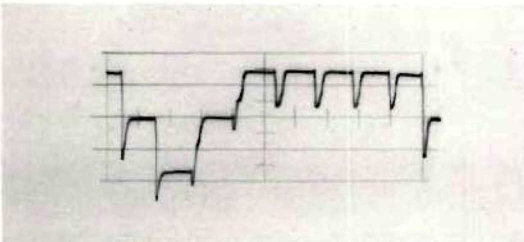
cycles per second. Primary use of the commutator is expected to require about 50,000 samples per second.

Figure 14(b) shows the large signal equivalent circuit of the transistor used as a switch (6,8). The transistor is represented as two diodes pointing toward each other in their easy conduction direction. In the absence of a switching pulse, the base of the transistor (the common connection of the two equivalent diodes) is biased to be positive with respect to both the collector and the emitter, by the voltage divider, R_1 and R_2. The diode, D_1, is forward biased, and the transistor is an "open" switch.

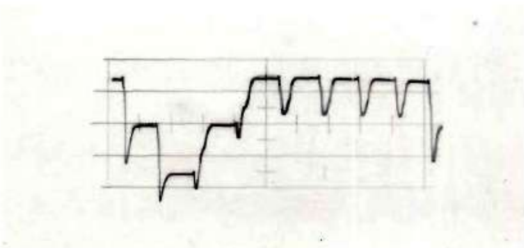
If a negative pulse, with an amplitude exceeding six volts, is applied at the point of connection of R_1 to D_1, the diode becomes back biased, thus disconnecting the sampling pulse from the transistor. The base voltage of the transistor becomes negative, and both the emitter-to-base and the collector-to-base diodes of the transistor are forward biased. The transistor then becomes a "closed" switch for the duration of the sampling pulse. The ratio of "open" to "closed" resistance of a typical transistor is of the order of 10^6.

The sampling pulse for the switch was derived from the positive going edge of the square wave synchronizing signal. The synchronizing pulses for the commutator were derived from the negative going wave fronts. The fast samples from the pulse stretcher were thus taken at the center of the samples emerging from the commutator, and transient effects were avoided.

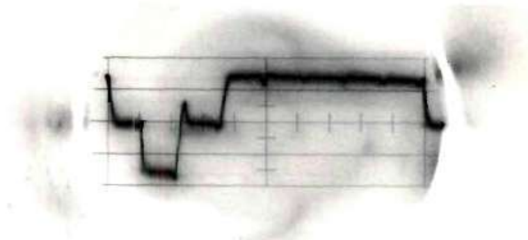
Figure 15 shows the output of the pulse stretcher, for sampling rates of 100,000 and 150,000 samples per second. For comparison, the



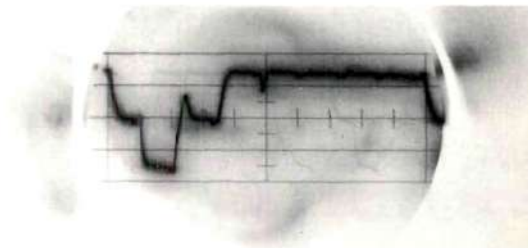
(a) 1.5, 3.0 and 1.5 Volts dc, 100 KC Sampling, with Fast Pulse Stretcher.



(b) 1.5, 3.0 and 1.5 Volts dc, 150 KC Sampling, with Fast Pulse Stretcher.



(c) 1.5, 3.0 and 1.5 Volts dc, 100 Kc Sampling, without Pulse Stretcher.



(d) 1.5, 3.0 and 1.5 Volts dc, 150 KC Sampling, without Pulse Stretcher.

Figure 15. Output Waveforms.

output of the commutator, without the pulse stretcher, is also shown. It will be noted that the transient when the pulse stretcher is used is shorter in duration, but larger in magnitude.

CHAPTER VII

CONCLUSIONS

The feasibility of a transistorized commutator for the analysis of voice signals, or signals having the approximate spectral limits of voice signals, has been demonstrated. Such a device can be constructed with general-purpose, point-contact diodes and low-cost transistors.

For voice signal analysis, a commutation speed of 40,000 samples per second is more than adequate. The commutator described here is capable of a rate of 100,000 samples per second.

Directions of design improvement have been implied. A moderate decrease in the duration of the switching transient in the output signal could be effected by use of the complementary symmetry multivibrator, but at the price of a larger number of transistors in the circuit. If improvement in output presentation alone is desired, without an increase in sampling rate, the pulse stretcher is an economical solution, because only the pulse stretcher need be a rapid (and expensive) element. If appreciably higher sampling rates are desired, however, the control multivibrators will require faster, more expensive transistors.

APPENDIX

To design a non-saturating bistable multivibrator by the method of Suran and Reibert (4,6) use Figure 9 and define:

ΔV_c = collector voltage swing desired,

ΔI_c = collector current swing desired,

v_{BE} = internal base-to-emitter drop of the conducting transistor,

$\cong$ 0.2 volt,

E_{bb} = supply voltage,

$A = \dfrac{\Delta V_c}{E_{bb}}$

I_{co} = reverse collector current, emitter open,

α_{bo} = common-base short-circuit current-transfer ratio, and

α_{eo} = common-emitter short-circuit current-transfer ratio.

Select a value for K such that

$$A > K \gg \frac{1}{A}\left|\frac{v_{BE}}{E_{bb}}\right| \text{, and}$$

$$(1 - A) > K \text{ .}$$

A value of K in the neighborhood of 0.2 is representative. Next define:

$$\psi = \frac{1}{K} \frac{v_{BE}}{E_{bb}} ,$$

and select a value of U such that

$$\alpha_{eo} \gg U > \frac{A}{(1 - \psi)(A - K)(1 + K - A)\alpha_{bo}} .$$

Calculate KX:

$$KX = \frac{U}{2} - \sqrt{(U/2)^2 - UA/(\alpha_{bo})(1 - \psi)} .$$

Make sure that

$$\Delta I_c \gg I_{co}(U - KX)(1 - \psi) ,$$

for the highest expected junction operating temperature. Calculate:

$$R_f = \frac{E_{bb}}{\Delta I_c} \left(K - \left| \frac{v_{BE}}{E_{bb}} \right| \right) ,$$

$$R_1 = XR_f ,$$

$$R_b = UR_f, \text{ and}$$

$$R_k = (U/K - U - X)R_f .$$

If f_M is the maximum pulse rate desired, C_k can be approximated:

$$C_k \cong U/(\pi f_M R_k K)(U + X) \;;$$

however, as noted in Chapter IV, if f_M is lower than the maximum pulse rate possible with the transistors being used, the value of C_k by the above approximation may be so large that the rise time of the output pulses will be too long. Selection of C_k by trial-and-error may be the more practical method.

If the circuit is triggered at the collectors, select the value for C_f such that

$$\frac{1}{f_M \pi R_f} > C_f \gg \frac{1}{R_f \omega_{\alpha b}} ,$$

where

$\omega_{\alpha b}$ = angular cutoff frequency, common-base circuit.

As an example, the multivibrator of Figure 2 will be computed. Given that

E_{bb} = 15 volts,

ΔV_c = 8 volts,

ΔI_c = 1.65 milliamperes,

v_{BE} = 0.2 volt,

α_{bo} = 0.95,

α_{eo} = 20, and

I_{co} = 50 microamperes,

then

$$A = V/E_{bb} = 0.53 ,$$

$$1 - A = 0.47, \text{ and}$$

$$v_{BE}/AE_{bb} = (0.2)/(0.53)(15) = 0.025 .$$

But

$$0.47 > K \gg 0.025 ,$$

therefore select

$$K = 0.2 .$$

Now

$$\psi = v_{BE}/KE_{bb} = 0.067 ,$$

and

$$\frac{A}{(1 - \psi)(A - K)(1 + K - A)\alpha_{bo}} = 2.43 ,$$

but

$$20 \gg U > 2.43 ,$$

therefore select

$$U = 4.0 .$$

Next compute

$$KX = 2.0 - \sqrt{(2.0)^2 - (4.0)(0.53)/(0.95)(0.93)} \; .$$
$$= 0.74 \; ,$$

then

$$X = 3.7 \; .$$

Compute

$$R_f = (15)(0.2 - 0.013)/(0.00165)$$
$$= 1700 \text{ ohms},$$
$$R_1 = (3.7)(1700)$$
$$= 6300 \text{ ohms} \; ,$$
$$R_o = (4.0)(1700)$$
$$= 6800 \text{ ohms, and}$$
$$R_k = (4.0/0.2 - 4.0 - 3.7)(1700)$$
$$= 21{,}000 \text{ ohms.}$$

And by trial-and-error

$$C_k = 230 \text{ micromicrofarads.}$$

The component values are not critical. Components with $\pm$ 10% tolerance were used in the commutator.

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